 By Morris Engelson, JMS Consulting

SOME LITTLE-KNOWN PROPERTIES ENABLE SPECTRUM ANALYZERS TO PROVIDE USEFUL RESULTS IN A SMALL FRACTION OF THE NORMAL MEASUREMENT TIME.

Speed up your spectrum-analyzer measurements

A spectrum analyzer's calibrated default sweep-speed setting assumes a specific application—the accurate measurement of the magnitude of a sinusoidal signal. The analyzer displays an "uncalibrated" message when it sweeps faster than the default setting. But just because the instrument isn't calibrated doesn't mean that it can't make the measurement. All that the uncalibrated message means is that the sine-wave-measurement error exceeds the limits that the instrument's calibrated-accuracy specification defines. Frequently, users are happy to accept a slight loss in accuracy to obtain a quicker result. After all, measurement errors exist even when the instrument is fully calibrated. Furthermore, many signals, such as those in pulsed radar and digitally modulated communications systems, aren't really sine waves. Yet, the analyzer's default settings treat all signals as if they were sine waves. Rather surprisingly, the analyzer loses little, if any, accuracy when it measures the vast majority of common signals at five to 10 times the calibrated speed.

Swept-frequency-analysis instruments, such as spectrum analyzers, are subject to an inherent measurement-time constraint. The signals must pass through a bandpass filter, whose output amplitude responds rather slowly to changes in input amplitude; the narrower the filter's passband, the more slowly the filter responds. If the signal frequency sweeps too rapidly through the filter's passband, the signal amplitude at the filter output is too low. In most cases, the minimum time required for making calibrated measurements is proportional to the reciprocal of 0.5 times the filter bandwidth (B) squared—the so-called 0.5 relationship. Measurement times of tens of minutes are not unusual when the resolution bandwidth is 10 Hz or less. Even at a bandwidth of several kilohertz, the measurement time is frequently many seconds.

This article discusses the relationship between sweep (measurement) time and accuracy and the impact of different signal types on accuracy. The article also provides procedures and examples that permit much faster measurements with, at most, a small loss in accuracy.

SWEEP-TIME LIMITATIONS

Reference 1 shows that the amplitude error for a sinusoidal signal displayed on a spectrum analyzer is equal to $\alpha=[1+(0.195(S/(B_R^2T))^2)]^{-1/4}$. In this equation, α is the ratio of the display amplitude to a normalized level of unity, S is the displayed frequency span, T is the sweep time, and B_R is the resolution bandwidth at the −3-dB points. Analyzer manufacturers usually choose an error default of 0.1 dB, which makes $\alpha=1/1.0116$. Choosing this amplitude level yields $S/(B_R^2T)=0.49$. Hence the generally accepted rule is that $S/T=0.5B_R^2$. For example, sweeping through a frequency span of 10 MHz with a 10-kHz resolution bandwidth calls for a sweep time of 200 msec to limit sine-wave errors to 0.1 dB.

The effect on measurement error is not linear with sweep time. For instance, in the previous example, a sweep time of 200 msec yields an α of 1.012, which is equivalent to a 0.1-dB error. Reducing the sweep time to 100 msec yields an α of 1.046, which represents nearly a 0.4-dB error. Still, given the spectrum

Figure 1

The spectrum of a continuous sine-wave signal under calibrated conditions and four times oversweep shows a somewhat reduced peak amplitude but has the correct shape.

analyzer's usual accuracy (really *in*accuracy), a 0.4-dB amplitude-measurement error may be perfectly acceptable in return for halving the measurement sweep time. **Table 1** provides the amplitude-error range for sinusoidal signals as a function of the dimensionless ratio $S/(B_R^2T)=k$.

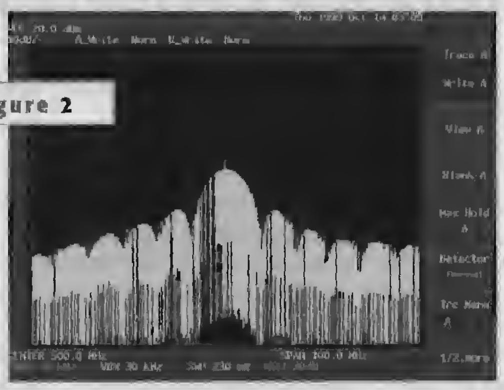
In the expression for k, the sweep time, T, is in the denominator. Therefore, reducing the sweep time from 3.9 sec to 0.5 sec—a factor of nearly eight—multiplies k by eight and introduces a 3-dB error. At k=5, which corresponds to a factor-of-10 reduction in sweep time, the error is 3.8 dB. Although 3 dB isn't particularly remarkable in this context, it has something of a special quality that people tend to choose. This choice explains a statement that you sometimes see in the literature—that you can increase the sweep speed to about 10 times the calibrated value before the amplitude error becomes really objectionable.

Obviously, you can use the above relationship to make any computation you choose. For example, the predicted amplitude error is 13.4 dB at a sweep-time reduction of 100 times. But even if you could accept a 13-dB error, such a speed-up would be impractical because several limitations prevent such a large reduction in the measurement time. The maximum practical sweep-time reduction is a factor of approximately 20, which produces a 6.6-dB error. A good rule to follow is that a five-times reduction is almost always possible, 10 times is usually possible, and 20 times is sometimes possible. There are rare situations in which time is all that counts. Usually though, time is important—but not at the expense of all else. In such cases, the so-called optimum-sweep-time relationship becomes useful.

OPTIMUM SWEEP TIME

Sweep time and resolution bandwidth are independent of each other. Hence, you can take a different view of the relationships among sweep width, sweep time, and resolution. What happens if you fix the frequency span and sweep time and vary the bandwidth? A narrower bandwidth violates the 0.5 relationship and causes a loss in amplitude. But there is also an impact on the displayed, or apparent, bandwidth. At relatively low sweep speeds where the 0.5 relationship holds, the displayed or dynamic bandwidth, essentially equals the true, or static, bandwidth. But the dynamic bandwidth increases as you reduce the sweep time. Is there an optimum value of static bandwidth that yields the narrowest displayed, dynamic bandwidth? Yes, and for this combination of bandwidth and span, the sweep time is optimum. The optimum bandwidth, B_O, is related to sweep time and span by a factor of 2.27 (**Reference 2**). This setting causes a 1.5-dB error in the measured amplitude.

Figure 2



The spectrum of a pulsed sine-wave signal under calibrated conditions and 11.5 times oversweep shows no amplitude reduction but a slight frequency shift, compared with the spectrum obtained at the normal (slower) sweep speed.

The result is not intuitively obvious. Why does this setting give the narrowest displayed bandwidth? Why can't you get something narrower? The explanation is that a wider static bandwidth obviously provides a wider displayed dynamic bandwidth. Both static and dynamic bandwidths are essentially equal when the 0.5 relationship applies. However, when the 0.5 relationship no longer holds, a narrower static bandwidth results in an increase in dynamic bandwidth. The rate of widening of the dynamic bandwidth is not linear with the change in static bandwidth, and the increase in dynamic width eventually exceeds the decrease in the static value. The crossover point establishes the optimum bandwidth.

If you hold the sweep time constant, achieving the optimum bandwidth requires a slightly greater than 2-to-1 reduction in bandwidth—$(2.27/0.5)^{1/2}=2.1$. The noise level is proportional to the bandwidth, so halving the bandwidth reduces the noise level by 3 dB. Hence, the noise drops by a bit more than 3 dB, the signal level drops by about 1.5 dB, and the SNR improves. Therefore, if you wish to sweep faster than usual, the ideal speedup is 2.27/0.5=4.5 times the preset calibrated value. This ratio provides the narrowest, or best, signal resolution and also produces the best SNR.

Yet, even a 1.5-dB measurement error can be significant. Surely, the situations in which such an error is acceptable must be rare. Not so. The reason is that fully half of spectrum-analysis measurements involve relative, rather than absolute, amplitude determination. Amplitude, frequency, and many other modulation measurements, as well as many spurious-response measurements, involve determining dBc, that is, signal-to-carrier ratio. In such measurements, the absolute amplitude error is of no consequence because both the carrier and the sidebands are similarly reduced in level, and the decibel difference remains unchanged. Another reason is that this error is systematic, rather than random. You can calculate a correction factor and adjust the final result accordingly. Hence, this procedure could enjoy much wider use—if more engineers knew about it.

The traces in **Figure 1**, taken from an Advantest 3132 spectrum analyzer, illustrates this idea. The two traces differ by 1.5 dB at a vertical setting of 2 dB per division. The span is 200 kHz, the bandwidth is 1 kHz, and the sweep time is 100 msec—all shown on the lower edge readout. This combination is not calibrated, as shown by the "uncal" message at the lower right. These settings result in the smaller trace, slightly to the right of center. The larger trace, to the left, was obtained at a calibrated sweep time of 400 msec. As required by the theory, the calibrated, S/B^2T setting is $2\times10^5/10^6\times$

TABLE 1—A SPECTRUM ANALYZER'S SINE-WAVE ERROR VERSUS A FUNCTION PROPORTIONAL TO THE RECIPROCAL OF THE RESOLUTION BANDWIDTH SQUARED

k	0.5	0.7	0.9	1.1	1.3	1.5	1.7	1.9	2.1	2.3	2.5	2.7	2.9	3.1	3.3	3.5	3.7	3.9
dB	0.1	0.2	0.32	0.46	0.62	0.79	0.97	1.16	1.35	1.54	1.73	1.92	2.11	2.29	2.47	2.65	2.82	2.99

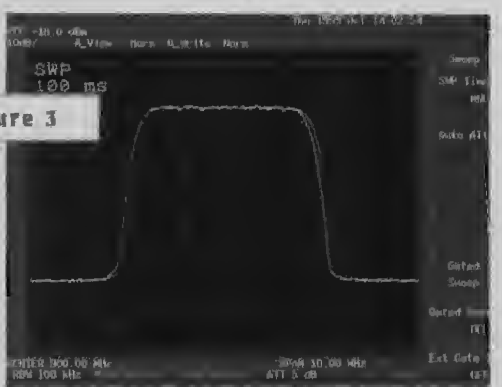
0.4=0.5. The sweep-time decrease is 400/100=4 times, for a value k=0.5× 4=2. Note that the four-times sweep-time reduction is the closest time setting in a 1, 2, 5 sequence to the previously calculated optimum value of 4.5. The theoretical impact on accuracy should be 1.3 dB. The measured value is slightly more than that, in line with the fact that real filters are somewhat less forgiving than a theoretically perfect filter would be.

NONSINUSOIDAL SIGNALS

You can oversweep nonsinusoidal signals (that is, sweep them at a speed greater than the fastest calibrated value) to an even greater degree than you can oversweep sine waves. As with sine waves, faster sweeps do not affect measurements of nonsinusoidal signals' relative amplitudes. However, most absolute-amplitude results are also unaffected. Hence, most nonsinusoidal signals are good candidates for oversweeping. The following two illustrations explain why.

Figure 2 shows the frequency spectrum of a pulsed-carrier signal. The span is 100 MHz, the bandwidth is 30 kHz, and the sweep time is 230 msec. All of these values appear on the lower line of the readout. The sweep relationship yields a value of 0.48, so the result is calibrated, as shown by the absence of an "uncal" message. The second trace, whose readout is not shown, is overlayed on the first trace. The second trace is not recognizable in the dense distributed spectrum because it is nearly identical to the first trace. In the **figure**, the sweep time is 20 msec, for a sweep-time speedup of more than 10-to-1. A sine-wave signal would show an amplitude difference of nearly 4 dB under the same conditions. Why is the pulsed signal not affected?

Figure 3



At 20 times oversweep, the spectrum of this digitally modulated signal does not distort any important signal characteristics.

The lobe spacing of the spectrum displays as 10 MHz at one-division width. The inverse is a pulse width of 100 nsec. The filter is being subjected to a 100-nsec-wide pulse and responds in a transient mode. The response of a 30-kHz-wide filter occupies 30 kHz/100 MHz=3×10^{-4} of a screen width. This value corresponds to a sweep time of 20 msec×3×10^{-4}=6 μsec at the faster 20-msec full-screen sweep. Hence, the narrower 100-nsec pulse width determines the filter's amplitude response regardless of the sweep time. You would have to speed up the sweep by more than 100 times to make the sweep time approach the pulse duration. Other factors would interfere with the measure-

ment well before then. Hence, you can obtain pulsed-signal spectra 10 and sometimes more than 20 times as fast as the spectrum-analyzer calibration suggests.

Figure 3 shows two traces of a wideband code-division multiple-access (WCDMA) signal. The calibrated time setting is 2 sec for the trace that is slightly to the left. The second trace, shifted to the right, was obtained at an uncalibrated sweep time of 100 msec. The display shows an "uncal" reading. The only effect of the 20-times speedup is to shift the trace slightly to the right. The amplitude levels of the two traces are identical, and the occupied-bandwidth values are virtually the same, though one is shifted slightly in frequency. The spectrum analyzer's total-channel-power measurement shows an integrated power level of −17.4 dBm for the calibrated settings and not quite 0.1 dB less with a faster sweep.

The reason for the lack of sweep-time impact on this signal's displayed level is different from that for pulsed signals. The sweep-time effect manifests itself during the rising and falling edges of the spectrum display. The sine-wave-based spectrum is so narrow that there is just not enough time for it to reach an equilibrium state. The same effect occurs for the WCDMA signal, but there is time to get to a full-level equilibrium condition. The result shows as a shift of the signal to the right, but the shape and magnitude are unaffected.

MORE THAN JUST SWEEP TIME

Sweep time is just one of the factors that affect a spectrum analyzer's total measurement time. Various control and processing functions introduce significant dead time. Sometimes the sweep time dominates the measurement time, and sometimes the dead time does. Which one dominates depends on several factors (**Reference 3**). This article deals only with issues that relate to sweep time.

The discussion shows that you don't degrade amplitude-measurement accuracy when you sweep broadband spectra as much as 20 times as fast as the spectrum analyzer's fastest calibrated sweep. Broadband spectra are typical of pulsed and digitally modulated signals. With a similar sweep speedup, sinusoidal-signal spectra show a loss in level, but you can calculate and correct for it. Moreover, faster sweeps do not affect relative measurements. You can achieve sweep-speed increases of five to 10 times and, on occasion, even 20 times. The ideal sweep-speed increase is 4.5 times the calibrated speed. The higher speed yields optimum signal-resolving power and SNR.

REFERENCES

1. *Handbook of Spectrum Analyzer Techniques*, Polarad Electronics, 1955.

2. Engelson, Morris, *Modern Spectrum Analyzer Theory and Applications*, pg 93 to 95, JMS Consulting.

3. "Optimizing spectrum analyzer measurement speed," Agilent Technologies, Application Note 1318.

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V_{CE} BREAKDOWN IS THE MAJOR CAUSE OF SEMICONDUCTOR-DEVICE FAILURE IN INDUCTIVE-LOAD CIRCUITS, BUT YOU CAN TAKE NUMEROUS STEPS TO PROTECT DEVICES FROM EXCESSIVE STRESS.

Techniques minimize switching-device failures in inductive circuits

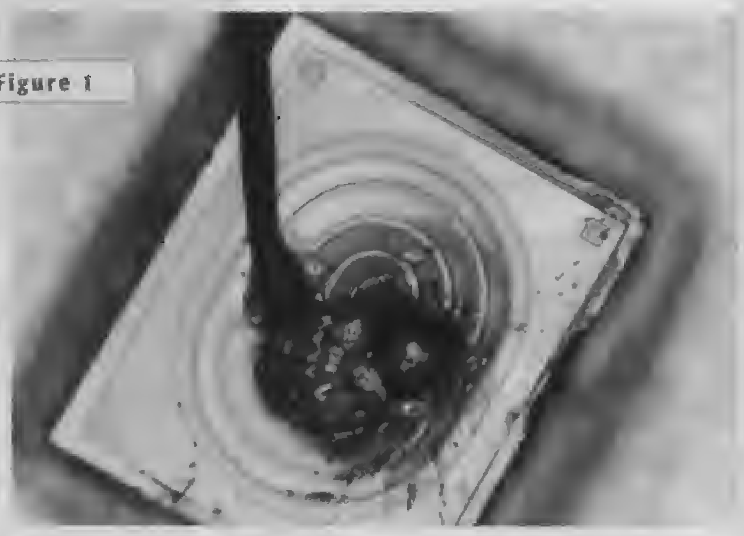
ALL ELECTRONIC CIRCUITS GENERALLY use inductive elements, such as transformers, coils, or relays, and some applications, such as stepper motors, also have inductive loads. Unfortunately, the failure of switching devices in inductive circuits is a frequent problem and causes system failure. Switching-transistor and MOSFET failure is common in motor drivers, switched-mode power supplies, and similar electronic systems. The causes for the switch failures can be intrinsic or extrinsic. In an intrinsic failure, the device itself may be faulty, it may have undergone degradation, or it may have a latent defect that initial device testing did not detect. Extrinsic causes of failure include simply misapplying the device or making it operate beyond its safe-operating-area characteristic during normal operation of the system. In this case, the device is bound to fail due to excessive stress. The stress can be electrical (**Figure 1**) or thermal overstress.

Minimizing switching-device failures in inductive circuits takes a multipronged approach. First, practice good design and construction techniques for transformers and other inductive elements in the circuit to reduce parasitics, such as leakage inductance and parasitic capacitance. Second, place protection networks or devices across the switching devices to protect against high-voltage breakdown. Third, use the appropriate switching device with a breakdown-voltage rating that's appropriate for the expected voltage levels in the circuit. You should derate by approximately 75% the breakdown-voltage rating that the data sheet specifies. You should specify acceptable leakage-inductance levels for all transformers and reject transformers that do not meet the specification during incoming inspection so that such defective components do not later cause field failures. Because the reliability of any system depends on the reliability of every component in the system, paying attention to minor parameter details of transformers and switching devices greatly improves overall system reliability by reducing component failures.

V_{CE} BREAKDOWN CAUSES MOST FAILURES

V_{CE} breakdown is the major cause of failure of semiconductor devices in circuits with inductive loads. Various reasons exist for the occurrence of this kind of failure. Whenever the voltage across the col-

Figure 1

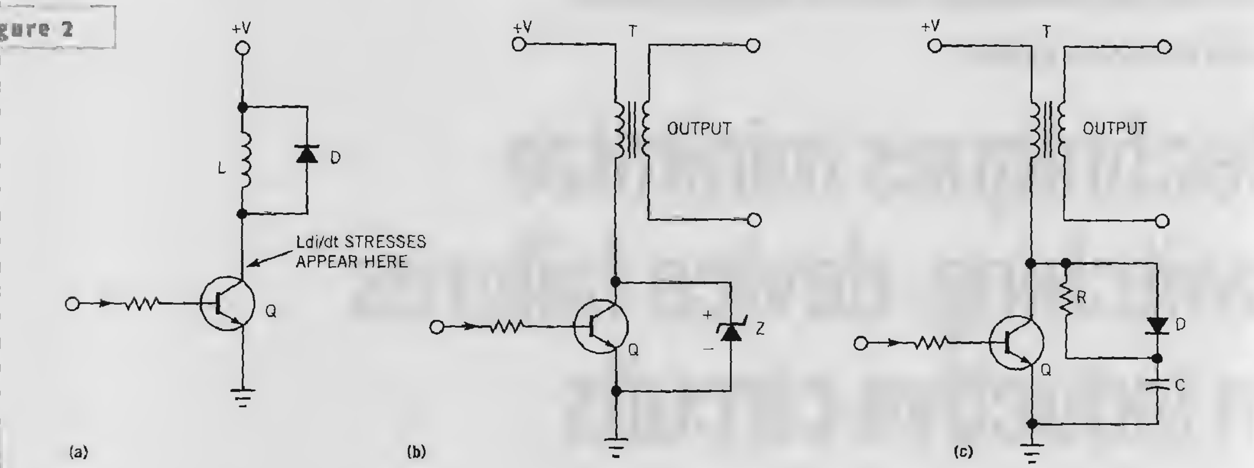


Electrical overstress severely damages a bipolar-junction transistor that operates in an inductive-load circuit—in this case, a switched-mode power supply.

lector-emitter terminals of the transistor exceeds its rating, the device is likely to fail. Although designers select devices depending on the application requirement and the expected voltage excursions in the circuit are likely to be within the V_{CE} rating, other elements in the circuit can cause the V_{CE} voltage to exceed the device's maximum limits.

Consider a basic transistor circuit in which the transistor acts as a switch, turning a motor winding on and off (**Figure 2a**). In this circuit, the switching of the current in the inductive-motor winding subjects the switching de-

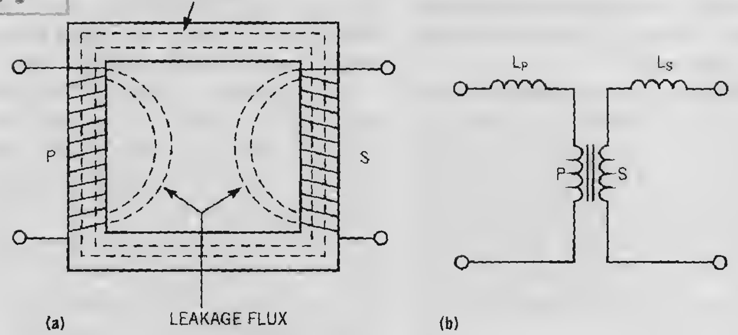
Figure 2



A diode protects the transistor from Ldi/dt switching spikes (a). A zener diode protects Q against V_{CE} breakdown (b). A snubber network protects Q from V_{ce} breakdown (c).

vice, Q, to stress. The stress arises due to the Ldi/dt voltage that the abrupt changes in the current through the motor winding generate due to the fast switching action. This stress manifests itself in the form of a voltage, which the circuit impresses across the collector-emitter terminals of the transistor. If the amplitude of the voltage across the collector-emitter terminals of the transistor exceeds its V_{CE} rating, high-voltage breakdown of the collector-emitter junction of the transistor occurs. You cannot apply the full rating specified for V_{CE} or any other parameter to a transistor. You have to derate the specified ratings depending on the application's reliability requirements and the operating temperature. Generally, you should derate the V_{CE} rating by 75%. Therefore, if a transistor is rated to withstand a maximum of 100V between its collector and emitter terminals as per data-sheet specifications, its V_{CE} rating is only 75V. In any application, the V_{CE} of this transistor should not exceed 75V. The lower the applied voltage between the collector-emitter terminals during normal operation of the system, the better the reliability of the system.

Figure 3



Leakage flux through air (a) causes leakage inductances L_P and L_S in the primary and secondary windings, respectively, of a transformer (b).

The first way to ensure the reliability of the system is to select a switching device that has a sufficient tolerance in voltage ratings, especially V_{CE} breakdown voltage, depending on the application requirement. Then, apply the appropriate derating, such as 75% for V_{CE}. You can also protect a transistor in an inductive load circuit from V_{CE} breakdown by connecting a diode across the inductive load

(**Figure 2a**), connecting a zener diode across the collector-emitter terminals of the transistor (**Figure 2b**), or connecting a snubber network across the switching device (**Figure 2c**).

In **Figure 2a**, Diode D provides a path for the current when the transistor turns off so that the decaying currents feed back to the power supply and the Ldi/dt spike does not cause stress across the collector-emitter terminals. This so-called freewheeling diode provides a free path for the decaying inductive currents when the switch is off.

In **Figure 2b**, the zener diode should have a breakdown voltage that's less than the transistor's V_{CE} breakdown rating so that the zener diode breaks down and conducts before a larger than normal voltage can damage the transistor. During normal operation, the zener diode does not conduct but rather breaks down when a larger than normal voltage appears at the transistor's collector. This zener diode, a so-called transient-suppressor diode, should have a fast response. Also, the diode's power-handling capability should be high enough to withstand the stress when the diode clamps a high voltage and when the instantaneous power dissipation is high.

In **Figure 2c**, the snubber network consists of a resistor and a capacitor connected in series across the transistor with a diode connected across the resistor. This network prevents the circuit from exceeding the reverse-bias safe-operating-area characteristics of the device. The snubber network reduces the stress on the transistor during on/off cycles and reduces EMI problems by reducing the dv/dt at the collector. You should select the snubber components so that V_{CE} breakdown of the transistor does not occur when no signal is present at the base of the transistor.

You can use the following guidelines, which stem from simple and familiar equations, to calculate the snubber-component values. Referring to **Figure 2c**, the fundamental capacitor-charge equation gives the amount of charge that the snubber capacitor stores, where Q is the charge, V is the voltage across the capacitor, and C is the capacitance of the capacitor:

$$q = Cv.$$

Differentiating with respect to time, t, gives the current, or the rate of change of charge:

$$\frac{dq}{dt} = C\frac{dv}{dt}.$$

Assuming 50% on and off times, the average current through the capacitor is I/2, where I is the full current. Thus,

$$C = \frac{I/2}{dv/dt} = \frac{I\,dt}{2\,dv}.$$

Over a full cycle, dt represents the fall time of the transistor's collector current. Remember that current flows through the capacitor only when the transistor is off. Thus, integrating the equation for C over a full cycle gives

$$2C\int dv = I\int dt.$$

Therefore, the value of snubber capacitor C is as follows:

$$C = \frac{\frac{I}{2} \bullet t}{0.75\ V_{CEO}} = \frac{I \bullet t}{1.5\ V_{CEO}}.$$

In this case, the voltage across the capacitor is approximately one diode drop lower than the voltage at the collector, and the equation includes a derating factor of 0.75 for the V_{CEO} rating of the transistor. When off, the transistor dissipates the energy that the capacitor stores. This energy is:

$$E = \frac{1}{2} CV^2.$$

In other words, the power dissipated in the transistor when it is off is:

$$P = \frac{1}{2} C(0.75\ V_{CEO})^2 f,$$

where f is the switching frequency of the transistor.

The RC time constant should be less than half of the on time of the transistor so that the snubber network is ready to charge when the transistor is off. Using this rule, you can calculate snubber resistor R as

$$R = \frac{1}{2} \bullet \frac{t_{ON}}{C}.$$

PARASITICS ALSO CAUSE PROBLEMS

In addition to V_{CE} breakdown, other problems, which commonly occur in switched-mode power supplies and other transformer-driven circuits, result from transformer parasitics. Leakage inductance and parasitic capacitance in the transformer are destructive parasitics that affect the reliable operation of the switching devices. A bad transformer can cause large power losses in a power supply due to parasitics. Normally, magnetic lines of force follow the magnetic circuit through the core between the primary and secondary windings. If the magnetic coupling between the primary and secondary windings is poor, magnetic lines of flux take other paths and complete the magnetic circuit through the air between the windings (**Figure 3a**). This leakage flux gives rise to leakage inductance in the primary and secondary windings, which manifests itself as small series inductances in the primary and secondary windings (**Figure 3b**). Similarly, a parasitic capacitance exists between the primary and secondary windings of a transformer. Also, capacitance exists between the turns of the windings and between the layers in a winding. In combination with leakage inductances, the stray and other parasitic capacitances can form resonant circuits, which can cause voltage spikes in the output circuits. You can measure the leakage inductance by shorting the secondary winding and measuring the inductance of the primary winding at the operating frequency of the transformer. Leakage inductance depends on the construction of the transformer but is not highly dependent on the frequency.

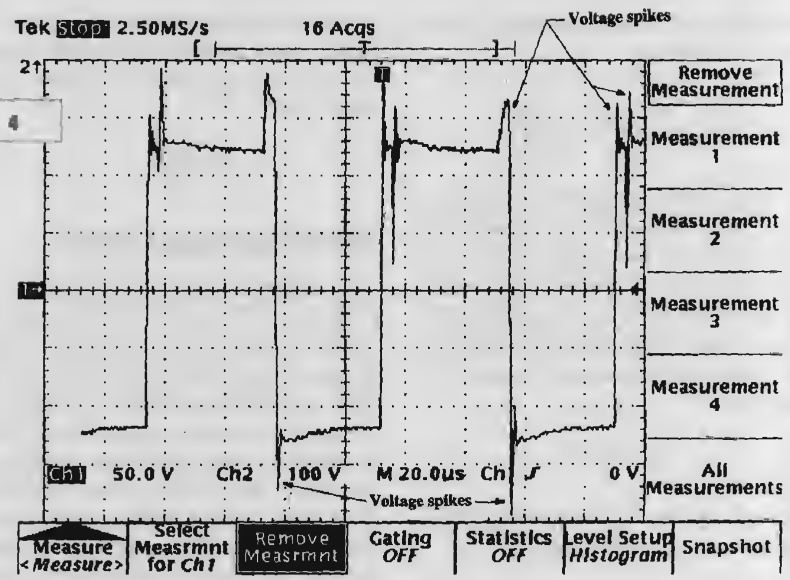
These leakage inductances can cause large output spikes, which in turn can

circuit failure. For example, in one recent case, a switching transistor in a transformer-based power-supply circuit was often failing. Large voltage spikes were observable at the output and the collector of the transistor, and the amplitude of the spikes was far above the normal switching waveform (**Figure 4**). Analysis of the mode of failure of the transistor revealed that the device had failed due to electrical overstress from collector-emitter breakdown (**Figure 1**). The amplitude of the voltage spikes exceeded the V_{CE} rating of the transistor, and, hence, the device had failed. Further analysis traced the cause of these voltage spikes to high leakage inductances in the transformer windings due to poor transformer construction and winding technique. Improving the transformer winding technique to reduce the leakage inductance corrected the problem.

Suitable transformer construction techniques include interspersing the primary and secondary windings to cover the whole bobbin and to ensure close proximity between the windings and placing an electrostatic screen between the primary and secondary windings.

Interspersing the windings and ensuring close proximity reduces flux leakage and results in better magnetic coupling between the windings. The key to reduc-

Figure 4



Voltage spikes are observable at the transition points of the collector waveform of the transistor.

ing leakage inductance is maximizing magnetic coupling between the primary and secondary windings. In the interleaved winding technique, you place the secondary winding between two layers of primary winding. In other words, you wind the primary winding in two parts that lie on either side of the secondary winding. Another technique for achieving close magnetic coupling between the primary and secondary is to twist the wires together before winding the transformer; this winding technique is called bifilar winding.

An electrostatic screen generally consists of a copper foil wide enough to cover the full winding area; it connects to ground through a connection at the center of the windings. You can ensure that the two ends of the foil do not touch by placing an insulating material before the screen ends overlap. If the ends are not insulated, the shorted copper foil will act like a shorted single-turn secondary winding, and overheating may destroy this foil due to the excessive current in this shorted foil, or "secondary." You ground the center point of the foil to cancel out the inductive effects due to the induced currents in the screen, which now flow in opposite directions to ground.□

AN ELECTROSTATIC SCREEN GENERALLY CONSISTS OF A COPPER FOIL WIDE ENOUGH TO COVER THE FULL WINDING AREA.

References

1. Lowden, Eric, *Practical Transformer Design Handbook*, Howard W Sams & Co, 1985.

2. Billings, Keith H, *Switchmode Power Supply Handbook*, McGraw-Hill Inc, 1989.

3. Brown, Marty, *Power Supply Cookbook*, Butterworth-Heinemann, 1994.

4. Brown, Marty, *Practical Switching Power Supply Design*, Academic Press, 1990.

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